

Pseudorandom noise code-based technique for thin cloud discrimination with CO₂ and O₂ absorption measurements

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Abstract

NASA Langley Research Center is working on a continuous wave (CW) laser based remote sensing scheme for the detection of CO₂ and O₂ from space based platforms suitable for ACTIVE SENSING OF CO₂ EMISSIONS OVER NIGHTS, DAYS, AND SEASONS (ASCENDS) mission. ASCENDS is a future space-based mission to determine the global distribution of sources and sinks of atmospheric carbon dioxide (CO₂). A unique, multi-frequency, intensity modulated CW (IMCW) laser absorption spectrometer (LAS) operating at 1.57 micron for CO₂ sensing has been developed. Effective aerosol and cloud discrimination techniques are being investigated in order to determine concentration values with accuracies less than 0.3%. In this paper, we discuss the demonstration of a pseudo noise (PN) code based technique for cloud and aerosol discrimination applications. The possibility of using maximum length (ML)-sequences for range and absorption measurements is investigated. A simple model for accomplishing this objective is formulated, Proof-of-concept experiments carried out using SONAR based LIDAR simulator that was built using simple audio hardware provided promising results for extension into optical wavelengths.

Keywords: ASCENDS, CO₂ sensing, O₂ sensing, PN codes, CW lidar

INTRODUCTION

The National Research Council's (NRC) Decadal Survey (DS) of Earth Science and Applications from Space has identified the Active Sensing of CO₂ Emissions over Nights, Days, and Seasons (ASCENDS) as an important Tier II space-based atmospheric science mission. The CO₂ mixing ratio needs to be measured to a precision of 0.5 percent of background or better (slightly less than 2 ppm) at 100-km horizontal resolution overland and 200-km resolution over oceans. To meet this goal, the ASCENDS mission requires simultaneous laser remote sensing of CO₂ and O₂ in order to convert CO₂ column number densities to average column CO₂ mixing ratios (XCO₂). As such, the CO₂ column number density and the O₂ column number density will be utilized to derive the average XCO₂ column. The anticipated benefits of ASCENDS as discussed in the decadal survey include (a) quantification of global spatial distribution of atmospheric CO₂ on scales of weather models as well as terrestrial and oceanic sources and sinks of CO₂ during day/night over all seasons, and (b) establish a scientific basis for future projections of CO₂ sources and sinks through data-driven enhancements of Earth-system process modeling.

Accordingly, NASA Langley Research Center (LaRC), with its partners, is working on a CW laser absorption spectrometer based remote sensing scheme operating in the 1.57 micron spectral band for the detection of CO₂ and 1.26 micron spectral band for the detection of O₂. For concentration determination, differential absorption lidar (DIAL) scheme at a selected transition for each gas is adopted. Hence, two are more wavelength at a known transition are utilized. The 1.5 micron spectral band lying in the telecom region was chosen for CO₂ detection due to spectroscopic properties combined with architectural advantages. The selection of 1.26 micron band for O₂ detection that is close to 1.57 micron spectral band instead of 0.76 micron spectral band is anticipated to provide processing advantages. Within these spectral bands, an optimal transition that is least sensitive to environmental parameters for each of these trace gases is selected. With limitations on CW laser powers, the remote sensing systems may have to operate in the photon starved regions. However, an Intensity modulated continuous wave (IMCW) technique is adopted to achieve highest possible SNR for processing return signals to minimize the influence of various noise sources within the filter bandwidths for each channel.

Unlike pulsed laser based DIAL systems, CW laser based remote sensing systems are affected by cloud and aerosol interferences. The reason for this is because the optical depth, which is a measure opaqueness of the atmosphere at a particular optical path length, is a function of the distance to the target in addition to the concentration of whatever chemical species one is trying to measure at the wavelength of the laser. One could use an additional laser altimeter to measure distance and that has been done in the past with some degree of success. However, in situations where there are thin clouds the distance one uses in the optical depth calculation is ambiguous because we have a return signal from the ground and thin cloud at the same time. The technique presented here gives one the ability to discriminate between the return from a cloud or the ground in a natural way by calculating a return profile as a function of distance in a manner analogous to pulse lidar. In essence, one uses a modulation scheme in combination with signal processing to convert that modulation pattern into a pulse. One laser wavelength is tuned to an absorption line of CO₂ which is the on-line wavelength, for instance, while the second laser wavelength is tuned to a wavelength just off that wavelength, which we call the off-line wavelength. One separates the two wavelengths by modulating each wavelength with an orthogonal sequence because a photo detector is not able to distinguish between such close wavelengths. Signal processing is used to separate out each wavelength and obtain a pulse profile for each wavelength. By taking the ratio of the pulse amplitudes for each wavelength we may calculate the CO₂ concentration from the optical depth and the distance. Accordingly, various realizable aerosol and cloud discrimination techniques are being investigated for determining required accuracies less than 0.5%. In this paper, the adaptation of a Pseudorandom Noise (PN) code based technique to discriminate returns from clouds, aerosols and ground is discussed¹. This is followed by proof-of-concept experiments carried out using acoustic frequencies.

Resolution and Range

The minimum resolution is given by

$$r = \frac{c}{2B_r} \quad (1)$$

where c is the speed of light and B_R is the bit rate of the PN code. The maximum range is given by

$$R = N r \quad (2)$$

where N is the code length. For example, for a PN code of length 255 bits and a bit rate of 50 kHz, the maximum range would be 764 km, which is the range we would need for a satellite. However, there are advantages to making the code longer as discussed in Section below.

PN CODES

PN codes are well known in the communications devices for their use in encoding communication channels such as in spread spectrum communications. One of the simplest PN codes is the m-sequence. Furthermore, maximum length ml-sequences are basis of many other PN codes. We use (ml) sequences. Besides radar and sonar, the ml-sequences have been used in Lidar applications for measuring range^{2,3}. Pulse lidars inherently provide range information. However, in the case of CW laser based remote sensing systems, range information on clouds and hard targets can be obtained by transmitting amplitude modulated PN code. Range profiles can be extracted by cross correlation between the reflected signal and the transmitted code. The correlation peaks of range profiles will indicate the location of scattering centers such as clouds and ground.

New codes that are orthogonal to each other over a limited operational range can be generated. *This gives us the ability to design a multichannel DIAL systems. As such, it allows one to measure relative absorptions of two or more laser wavelengths at a particular range to detect specific atmospheric species.* Besides m-sequences^{4,5} other codes are also useful for ranging⁶. However, the use of time shifted PN codes to modulate multiple laser wavelengths for measuring absorption at specific ranges in order to discriminate the return of a cloud from the ground is a very effective and simple solution. For our application, the m-sequences are especially advantageous because they have very good autocorrelation properties.

The ML sequences can be represented in a number of different ways. One popular method is by using linear feedback registers with modulo 2 additions⁷ which is shown in Figure 1.

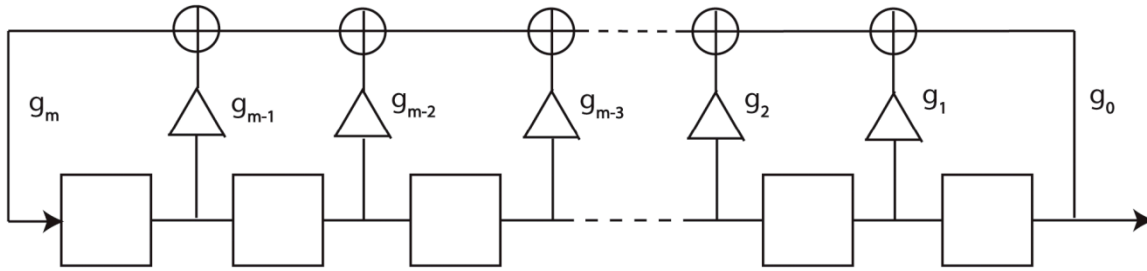


Figure 1. Shift register representation for m-sequence codes.

Another representation is using a generator polynomial

$$G = g_m x^m + g_{m-1} x^{m-1} + g_{m-2} x^{m-2} + \dots + g_2 x^2 + g_1 x^1 + g_0 \quad (3)$$

These sequences are 2^m-1 in length and only specific polynomials can be used for a particular length. For $m=8$ which corresponds to a code length of $2^8-1=255$, an allowed polynomial is,

$$G = x^8 + x^6 + x^5 + x^2 + 1 \quad (4)$$

In order to generate the individual bits, we can represent this by the recursion relation

$$X_{j+8} = X_j \oplus X_{j+2} \oplus X_{j+5} \oplus X_{j+6}, \quad (5)$$

where X is the maximum length sequence and j is the index. For the initial values of $X_0 - X_7$ we may choose any sequence of 0s and 1s we wish as the seed as long as they are not all 0 or all 1. For the seed (1,0,1,0,1,1,1,1) we generate the repeating sequence

(1, 0, 1, 0, 1, 1, 1, 1, 0, 0, 1, 1, 1, 0, 1, 0, 0, 0, 0, 1, 0, 1, 0, 1, 1, 0, 0, 1, 0, 1, 0, 0, 0, 1, 0, 1, 1, 0, 0, 0,
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1, 0, 1, 0, 0, 1, 1, 1, 0, 0, 0, 1, 1, 1, 1)

Correlation properties

As mentioned earlier, m-sequence codes have very good autocorrelation properties. The cross correlation between itself and a shifted version of itself is N (code length) when they are in sync and -1 if they are not. It is also understood that the codes are AC coupled before any correlation is done by changing all 0's in the code by -1 . There are a number of ways of computing the correlation. One is by computing a table of

$$R[A, B] = \sum_{m=1}^N A[m]B[m+n] \quad (6)$$

Note that some people define this differently by dividing by the length. This works but it requires something of the order of N^2 steps. A better and faster way is to use Fourier transforms. In this case,

$$R[A, B] = F^{-1} \left[F[A]^* F[B] \right] \quad (7)$$

where the $*$ indicates the complex conjugate and F is the FFT. This is superior because it only takes something of the order of $N \log N$ steps. As an example we take the case in the previous section and

compute the cross correlation between it and a code shifted 128 places to the right and the result is shown in Figure 2.

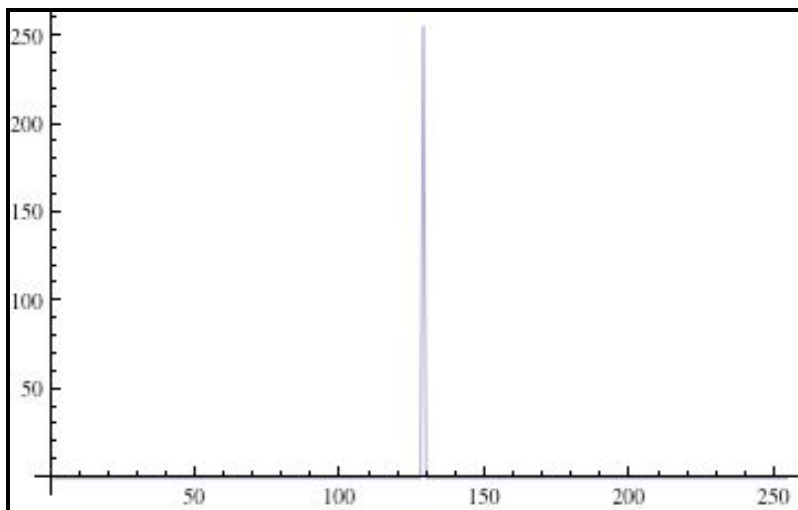


Figure 2. Cross correlation of code and shifted code computed with FFT.

Multiple channels

In order for this to be useful for atmospheric species detection it is advantageous to be able to separate out multiple transmitted wavelengths. Each wavelength is modulated with a different orthogonal code. We may take advantage of two facts to use m-sequences (for instance) to construct a system with two or more channels, where each channel represents a different wavelength that is modulated by a different code. Since m-sequences have very good autocorrelation properties shifted versions of them will cross correlate well except for a small DC component where they are uncorrelated and the big spike where they correlate. The second thing we have with this application is when the LIDAR is looking down from space there will be a final hard target (the ground) below which there will be no further returns possible. That opens up some interesting possibilities. If we know there will be no further returns past a certain distance we can size a system with twice the range we need and use that further range to add a second shifted version of that same code that is shifted by half the doubled range. To get twice the range we simply use a PN code twice as long while maintaining the same code bit rate. As long as there are no returns beyond that half range there will be no interference between the channels. The first half of the cross-correlation gives the range info for the first unshifted channel and the second half of the autocorrelation gives the range info for the shifted second channel and we can do that in a single step. By doubling the length of the code we also increase the SNR as we see from the previous section. We can make the code as long as we need to add more channels.

In the sample below we use a code length of 511 with a sample rate of 500000 samples/sec and a code bit rate of 50000 bits/sec. This gives us an over-sampling of $M=10$ and gives us a total range of $511 \cdot 3 \cdot 10^5 / (2 \cdot 50000) = 1530$ km. To generate the code we use a code with the generating function $G = x^9 + x^7 + x^5 + x^2 + 1$, a seed of (1,0,1,0,1,1,1,1,1) and a shifted version of that code shifted 256 places to the right. The result of that and 10x noise is shown in Figure 3.

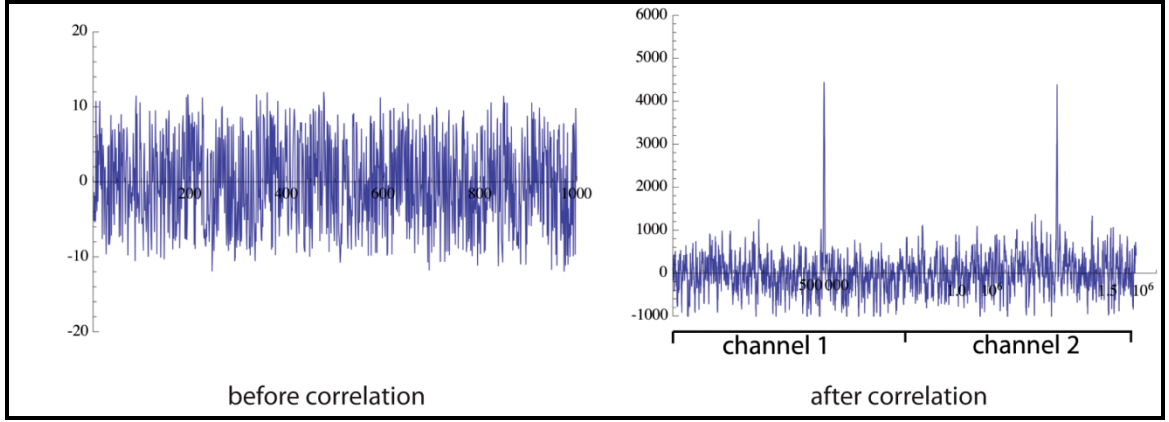


Figure 3. Example of using two codes in a double ranged system with two channels.

Signal-to-noise ratio estimates

We first consider white noise. It is possible to estimate the signal to noise ratio improvement before and after correlation. Let S_R be the sample rate and B_R be the PN code bit rate. Then number of samples per code bit is

$$M = \frac{S_R}{B_R} . \quad (8)$$

Oversampling the code by a factor of M increases the length of the code by the same factor so the code length becomes MN .

Lets suppose we have an AC coupled signal which we represent by

$$S = \eta + \alpha \, pn^\Delta \quad (9)$$

where η is the noise, α is the attenuation factor and

$$pn_i^\Delta = 2PN_{i+\Delta} - 1 \quad (10)$$

The term PN is the PN code represented by 1's and 0's and Δ represents the phase shift and pn contains 1's and -1's. The standard deviation of the noise we represent by

$$\sigma = \sqrt{\sum_{i=1}^{MN} \frac{(\eta_i - \eta_{ave})^2}{MN}} \quad (11)$$

where we have summed over the total length of the code. After performing the correlation calculation, we see that the signal strength increases by a factor of MN where they are correlated. We now ask the question what correlation does to the standard deviation of the noise. The correlation function of the total signal is,

$$R(j) = \sum_{i=1}^{MN} (\eta_i + \alpha p n_i^\Delta) p n_{i+j}^0 = \sum_{i=1}^{MN} \eta_i p n_{i+j}^0 + \alpha \sum_{i=1}^{MN} p n_i^\Delta p n_{i+j}^0 = \eta_j' + S_j' \quad (12)$$

The question now is to find the standard deviation of the transformed noise η' which we assume to have 0 mean (AC coupled). We first find the variance by

$$(\sigma')^2 = \langle (\eta')^2 \rangle = \frac{1}{MN} \sum_{j=1}^{MN} (\eta_j')^2 \quad (13)$$

To find the sum we write

$$(\eta_j')^2 = \sum_{m=1}^{MN} \eta_m p n_{m+j}^0 \sum_{n=1}^{MN} \eta_n p n_{n+j}^0 = \sum_{n,m} \eta_m \eta_n p n_{m+j}^0 p n_{n+j}^0 \quad (14)$$

so that

$$(\sigma')^2 = \frac{1}{MN} \sum_{j=1}^{MN} \sum_{n,m} \eta_m \eta_n p n_{m+j}^0 p n_{n+j}^0 \quad (15)$$

If the noise has 0 mean, the only part of the sum that will contribute is when $n=m$ so

$$(\sigma')^2 = \frac{1}{MN} \sum_{j=1}^{MN} \sum_m \eta_m^2 (p n_{m+j}^0)^2 = \frac{1}{MN} \sum_{j=1}^{MN} \sum_m \eta_m^2 = \frac{1}{MN} \sum_{j=1}^{MN} MN \sigma^2 = MN \sigma^2. \quad (16)$$

In the above since pn is either a 1 or -1, its square is always 1. The result is that

$$\sigma' = \sqrt{MN} \sigma. \quad (17)$$

The new correlated snr in terms of the original snr is

$$snr' = \frac{NM\alpha}{\sqrt{MN}\sigma} = \sqrt{MN} snr \quad (18)$$

Expressed differently in terms of previously defined variables we find

$$\frac{snr'}{snr} = \sqrt{MN} = \sqrt{\frac{samplerate}{codebitrate}} N \quad (19)$$

It should be stressed that this was derived assuming noise with 0 mean. If the DC offset is not subtracted from the signal before processing we would get a very different result because the PN code has a small DC offset for the simple reason it has an odd length. Alternatively one may subtract the offset from the code. The correlation properties would be similar and the resulting code would be orthogonal to any DC bias in the data.

Photo detector SNR

If the system is lidar we must use the SNR equation for photodiodes. As an example for two channel PN code modulation we use an on-line and a time shifted off-line code to modulate the on-line and off-line wavelengths. The SNR for such a signal is.

$$SNR = \frac{\sqrt{\langle i_s^2 \rangle}}{\sqrt{2q(\langle I_s \rangle + \langle I_N \rangle)}} \frac{1}{\sqrt{\Delta f}}. \quad (20)$$

Here I_s represents the signal current and I_N represents the equivalent noise current from all noise sources, whereas i is the AC coupled part of I . The variables q and Δf are the electron charge and noise bandwidth. For the received PN signal the SNR for each channel would be

$$SNR_n = \frac{\Re \sqrt{\langle p_{sn}^2 \rangle}}{\sqrt{2q(\Re \langle P_{s1} \rangle + \Re \langle P_{s2} \rangle + \Re \langle P_N \rangle)}} \frac{1}{\sqrt{\Delta f}} \quad (21)$$

Here $\langle P_{s1} \rangle$ and $\langle P_{s2} \rangle$ are the average on and off-line signal power, whereas $\langle P_N \rangle$ and \Re are the equivalent total noise power and responsivity. We use the RMS power in the numerator because only the AC coupled part of the signal is used in the processing. The variable p represents the AC coupled part of P . We first write

$$pn = 2PN - 1. \quad (22)$$

Since $\langle PN \rangle = (N + 1) / (2N)$, we find

$$P_{sn} = \langle P_{sn} \rangle \frac{N}{N + 1} (1 + pn). \quad (23)$$

The AC coupled part of P is

$$p_{Sn} = \langle P_{Sn} \rangle \frac{N}{N+1} \left(pn - \frac{1}{N} \right) \quad (24)$$

The $1/N$ term is the slight dc offset of pn . From this we find

$$\sqrt{\langle p_{Sn}^2 \rangle} = \langle P_{Sn} \rangle \frac{N}{N+1} \sqrt{\left\langle \left(pn - \frac{1}{N} \right)^2 \right\rangle} = \sqrt{\frac{N-1}{N+1}} \langle P_{Sn} \rangle. \quad (24)$$

Therefore,

$$SNR_n = \sqrt{\frac{N-1}{N+1}} \frac{\Re \langle P_{Sn} \rangle}{\sqrt{2q \left(\Re \langle P_{S1} \rangle + \Re \langle P_{S2} \rangle + \Re \langle P_{bg} \rangle \right)}} \frac{1}{\sqrt{\Delta f}}. \quad (25)$$

We may now use the results of Equation 18 to find the SNR after correlation. If we subtract the bias from the noise before correlation we find the new SNR after correlation is

$$\begin{aligned} SNR'_n &= \sqrt{\frac{N-1}{N+1}} \frac{M N \Re \langle P_{Sn} \rangle}{\sqrt{M N} \sqrt{2q \left(M N \Re \langle P_{Sn} \rangle + \Re \langle P_{Sn'} \rangle + \Re \langle P_N \rangle \right)}} \frac{1}{\sqrt{\Delta f}} \\ &= \sqrt{M N} \sqrt{\frac{N-1}{N+1}} \frac{\Re \langle P_{Sn} \rangle}{\sqrt{2q \left(M N \Re \langle P_{Sn} \rangle + \Re \langle P_{Sn'} \rangle + \Re \langle P_N \rangle \right)}} \frac{1}{\sqrt{\Delta f}}, \end{aligned} \quad (26)$$

where n and n' take on the values of 1 or 2 and $n \neq n'$. The extra MN term under the square root comes about because that part of the noise is correlated with the signal. This compares favorably with what has been reported elsewhere² if the differences in peak versus average power, RMS power, over sampling, and number of channels are taken into account.

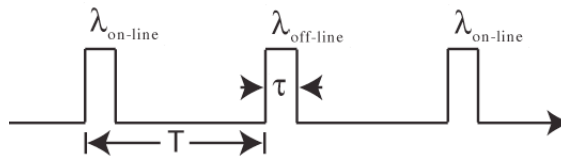


Figure 4. Pulse lidar system for on-line/off-line measurement

Comparison with pulse

For a pulse lidar system as shown in Figure 4 the SNR is

$$SNR_{pulse} = \frac{\Re P_{pulse Sn}}{\sqrt{2q(\Re P_{pulse Sn} + \Re \langle P_N \rangle)}} \frac{1}{\sqrt{\Delta f}} \quad (27)$$

for one complete on-line/off-line cycle. In order to match the resolution for pulse we have

$$\frac{c}{2B_r} = \frac{c \tau}{2} \Rightarrow B_r = \frac{1}{\tau} \quad (28)$$

If we make the sample times the same for a complete cycle we have $N = 2T B_r = 2T / \tau$. If we make the noise bandwidth the same and make the average signal power for the PN system the same as the pulse system multiplied by a factor of ρ then by Equation 26 the PN SNR is

$$\begin{aligned} SNR_{PN} &= \sqrt{M} \sqrt{\frac{N-1}{N+1}} \sqrt{\frac{2T}{\tau}} \frac{(\rho \tau / (2T)) \Re P_{pulse Sn}}{\sqrt{2q(\rho M \Re P_{pulse Sn} + (\rho \tau / (2T)) \Re P_{pulse Sn'} + \Re \langle P_N \rangle)}} \frac{1}{\sqrt{\Delta f}} \\ &= \sqrt{\rho} \sqrt{\frac{N-1}{N+1}} \sqrt{\frac{\tau}{2T}} \frac{\Re P_{pulse Sn}}{\sqrt{2q(\Re P_{pulse Sn} + (\tau / (2TM)) \Re P_{pulse Sn'} + \Re \langle P_N \rangle / (\rho M))}} \frac{1}{\sqrt{\Delta f}} \end{aligned} \quad (29)$$

For the same noise bandwidth and noise levels we find

$$\frac{SNR_{PN}}{SNR_{pulse}} = \sqrt{\rho} \sqrt{\frac{N-1}{N+1}} \sqrt{\frac{\tau}{2T}} \frac{\sqrt{P_{pulse Sn} + \langle P_N \rangle}}{\sqrt{P_{pulse Sn} + (\tau / (2TM)) P_{pulse Sn'} + \langle P_N \rangle / (\rho M)}} \quad (30)$$

In the strong signal limit we have

$$\frac{SNR_{PN}}{SNR_{pulse}} = \sqrt{\rho} \sqrt{\frac{N-1}{N+1}} \sqrt{\frac{\tau}{2T}} \sqrt{\frac{P_{pulse Sn}}{P_{pulse Sn} + (\tau / (2TM)) P_{pulse Sn'}}} \quad (31)$$

In the weak signal limit we find

$$\frac{SNR_{PN}}{SNR_{pulse}} = \rho \sqrt{M} \sqrt{\frac{N-1}{N+1}} \sqrt{\frac{\tau}{2T}} \quad (32)$$

The strong signal limit is a case we would not likely see since the detector shot noise will typically be less than other noise sources. As a result we are left with the weak signal limit. We may use this

to optimize the CW system for the best possible performance in terms of sample rate and average power level. Of course there are other considerations as well. In any real system one must consider the specific implementation when making such a comparison. One obvious advantage of the CW system is it can make a simultaneous on-line/off-line measurement whereas with the pulse system the measurement is sequential. This can potentially make the pulse system less accurate in determining the optical depth ratios.

EXPERIMENTAL RESULTS AND DISCUSSION

To test these concepts we implement the system shown in Figure 5 and 6 in audio^{8,9}. Because an audio speaker is not capable of reproducing the very low frequencies required by the PN code, one must first put the PN code on a carrier, then modulate that carrier with the PN code. Here the resolution and range are represented by Equations 2 and 3 except that c becomes the speed of sound in this case.

In the initial development, this system was first implemented in Mathematica⁸ then in Labview. In order to accomplish this, special Labview audio drivers written by Christian Zeitnitz¹⁰ were utilized. This increased both the performance and functionality of the Labview audio interface. A Samson Servo 201a audio amp was used for the audio amplifier. This had a bandwidth of 65 kHz and is designed for accurate sound reproduction. The audio card was an M-Audio 1814 with a 1394a interface. This card is capable of sample rates of up to 192 kHz. The software display is shown in Figure 7.

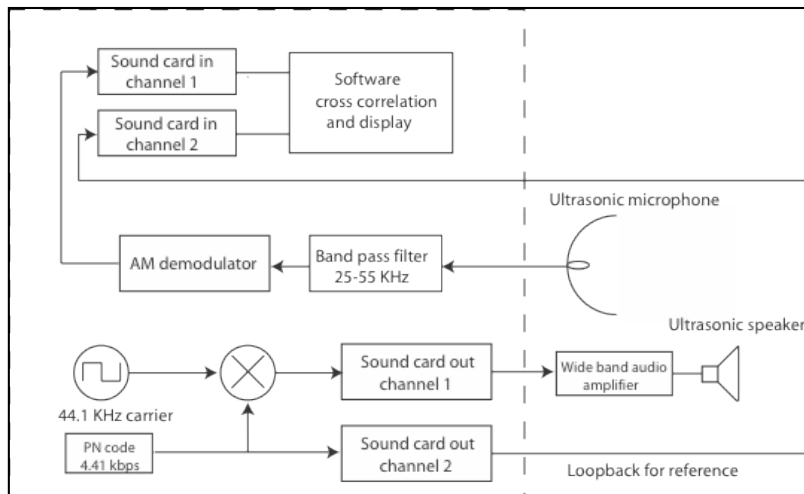


Figure 5. Audio implementation for single channel audio ranging system. All modulation/demodulation, filtering, and cross correlation functions were implemented in Labview software.

A dual channel implementation is also possible which could potentially be used for differential absorption measurements. This is analogous to the proposed LIDAR implementation and can potentially be used for differential absorption measurements.

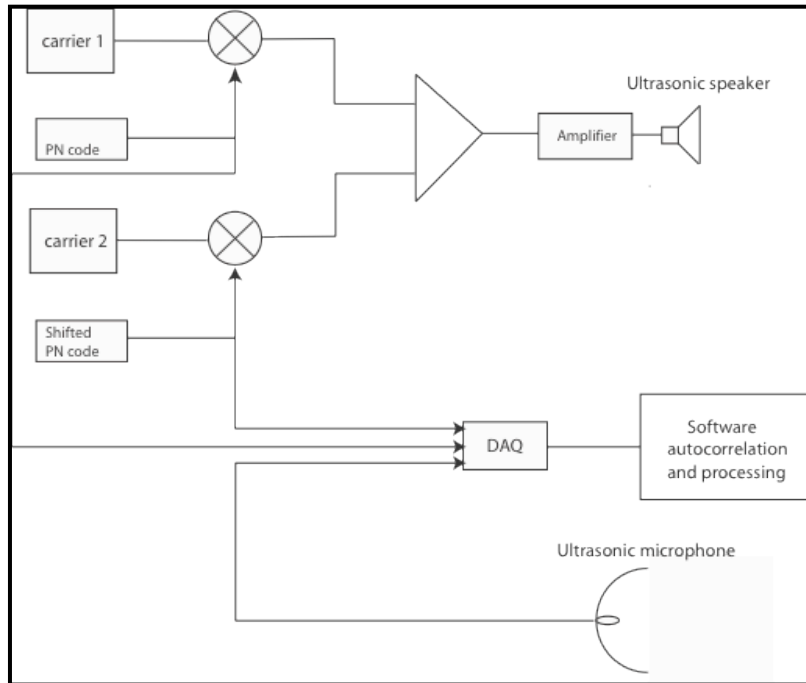


Figure 6. Dual channel implementation of audio ranging system can potentially be used for differential absorption measurements. This is analogous to how it could potentially be implemented in LIDAR.

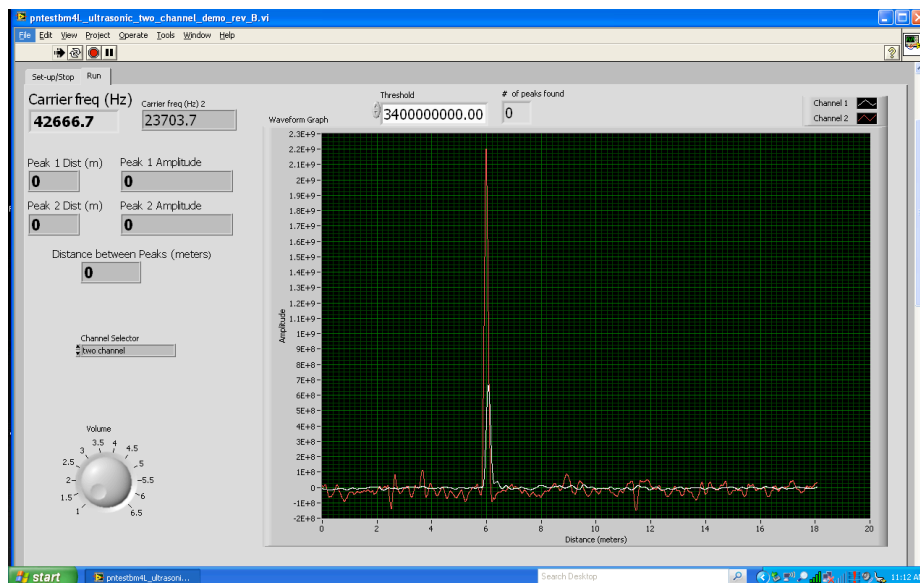


Figure 7. Labview interface and display for the dual channel implementation.

The proposition is to use two or more PN codes where each is shifted in time in a LIDAR implementation. The shifting of the codes is done such that there is equal spacing. For instance with two codes of length 511, the second is shifted to the right 256 places with respect to the first code. The shifting is done in such a way that as the end of the code is shifted past the boundary it wraps to the beginning. The maximum distance to the farthest target shall be 255 in that case. With four codes of length 511, each is shifted 128 places to the right of the preceding code. The maximum distance to the target in that case shall be 127. In this way one may have multiple channels for a given code length. The only restriction is that the LIDAR/RADAR/SONAR must be pointed at targets with a hard surface (such as the surface of the earth) no farther than the distance to the first code. Without that one will have range-wrapping issues. In addition to this one should either subtract the average from the code or from the data prior to processing. This helps eliminate noise/bias issues.

CONCLUSIONS

In this paper, a novel method based on PN codes to discriminate clouds and aerosols from ground returns suitable for CW laser based remote sensing systems for space based platforms is discussed. The intended use of this technique is in the ongoing for ACTIVE SENSING OF CO₂ EMISSIONS OVER NIGHTS, DAYS, AND SEASONS (ASCENDS) program effort at NASA Langley Research Center. It is shown that time-shifted PN codes would be valuable for obtaining range information of interfering clouds and aerosols so that returns from ground or topographic background can be discriminated during data retrieval processes. Proof-of-concept prototype operating at audio frequencies has been successfully demonstrated. Plans are underway for demonstration at optical frequencies.

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